Taken to its logical conclusion, this result suggests that the encoding of information onto the location burst is not efficient and the use of the location burst for data messages is therefore questionable[6] because the total data capacity could be doubled by using two separate systems. In this case, TIA's original analysis applies.

If a prime purpose of a location system is high data capacity, and information is encoded onto the location pulse, then the result is a system with poor processing gain and poor range, due to the desire to keep the code length short. This can be clearly seen in the Pinpoint "Array" system, an analysis of which can be found in Annex 1 of the "Reply Comments of MobileVision, L.P., July 29, 1993".

For best location capacity it is better to have as long a code length as possible[7]. If the prime purpose of the system is location, then doubling the bandwidth can increase the location capacity fourfold, provided the range is reduced in order to maintain the received signal above the threshold. In order to achieve high location and information capacity, the MobileVision system uses narrow band channels within the allocated band. Doubling the bandwidth then results in doubling the information capacity independent of the factors affecting location capacity.

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[6]In the MobileVision system, the information in the location burst is usually restricted to essential data, such as identification. It is understood that Teletrac does not encode any data into the location burst, whereas Pinpoint encodes all its data in the spread spectrum bursts.

[7]As the code length increases, the frequency accuracy demand becomes higher and also the aquisition time becomes longer. Therefore, there is a practical limit to the code length of about 1023.

2

Technical Note
Basic Relationships
concerning
Location using Direct Sequence Spread Spectrum

Summary

This Technical Note derives and discusses the basic relationships between the following elements of a direct sequence spread spectrum location system:

- duration of location pulse (location capacity)
- data capacity of the location burst
- processing gain
- jamming margin
- near-far-ratio
- range
- multipath errors

The effect of bandwidth and code length is further analyzed. It is shown, in the first example, that, if the code length is kept constant, the effect of doubling the bandwidth is to:

1) increase the system capacity by a factor of four,
2) increase the data capacity of the location burst by a factor of two,
2) reduce the range by 0.82.
3) double the variable multipath error,
and 4) halve the multipath bias error.

The significant gain in system capacity is offset by a reduction in range. The effect on multipath error tends to cancel out if both stationary and moving vehicles are considered. The 0.82 reduction in range equates to reducing the area of a cluster by 0.66.

In the second example, it is shown that the effect of doubling both the code length and the bandwidth is to:

1) increase the system capacity by a factor of four,
2) maintain the same location burst data capacity,
2) maintain the same range,
3) double the variable multipath error,
and 4) halve the multipath bias error.

In comparison with the first example the range is maintained at the expense of data capacity. From this it can be seen that the two requirements of location capacity and data capacity are somewhat in opposition. If the bandwidth is increased, by the use of a faster chipping rate, it is better to also increase the code length so as to maintain the range. This results, however, in no gain to the data capacity of the location burst. If the location burst is used for data transmissions, then it is necessary to keep the code length, L, constant. Therefore, as the chipping rate is increased, in order to increase the location capacity, the range reduces.

Taken to its logical conclusion, this result suggests that the encoding of information onto the location burst is not efficient.[1] and the use of the location burst for data messages is therefore questionable.

[1] It is for this reason that the MobileVision system uses narrow band channels within the allocated band. Doubling the bandwidth then results in doubling the data/voice capacity independent of the location capacity.

If a prime purpose of the system is high data capacity, and information is encoded onto the location pulse, then the range and processing gain will probably suffer. If the prime purpose of the system is location, then doubling the bandwidth is definitely more efficient in that the location capacity is increased fourfold.

In order ot achieve high gain and data capacity, it is necessary not to use the location pulse for the data transmission. For example, the MobileVision system uses narrow band channels, within the allocated bandwidth, for data and voice communications. Doubling the bandwidth then results in doubling the information capacity independent of the factors effecting the location capacity. Under these conditions, doubling the bandwidth undoubtably increases the efficiency of the system.

Glossary of Terms

A	amplitude
B	bandwidth
Bss	chipping frequency expressed as a bandwidth
D	distance, range
Dj	distance of jamming transmitter
Em	epoch error due to multipath
Fc	chipping rate
fd	doppler frequency
hb	base station antenna height
JM	jamming margin
L	length of spreading code
M	number of independent samples, number of epochs, number of data symbols
N	noise power
No	noise power per unit bandwidth
PG	processing gain
Pr	received power
Prj	received power of jamming signal
Pt	transmitted power
R	desired rms timing error
S	number of data symbols in unit time
$\left(S/N\right)_{out}$	signal to noise ratio at output of correlator
Tc	duration of chip
Tp	duration of location burst
Ts	duration of single spread spectrum code sequence, duration of epoch
V	vehicle speed
σ_t^2	variance of timing error
ϕ	phase constant
λ	signal wavelength

1. Relationship between Location Capacity and Bandwidth

In Appendix A to this Note, the variance of the timing error is shown to be:

$$\sigma_t^2 = \frac{1}{Bss^2 \left(\frac{S}{N}\right)_{out}} \tag{1}$$

where $\frac{1}{Bss} = Tc = 1$ chip duration

If M is the number of independent samples, then the variance of the average is:

$$\sigma_{ave}^2 = \frac{\sigma_t^2}{M} \tag{2}$$

Let the desired rms timing error be R. Then:

$$M = \frac{\sigma_t^2}{R^2} \tag{3}$$

Thus, $$M = \frac{1}{R^2 \, Bss^2 \left(\frac{S}{N}\right)_{out}} \tag{4}$$

Now the length of a pulse, $Tp = M.Ts$ where Ts is the duration of a single spread spectrum sequence.

Therefore $$Tp = \frac{Ts}{R^2 \, Bss^2 \left(\frac{S}{N}\right)_{out}} \tag{5}$$

Now $$Ts = \frac{L}{Fc} \tag{6}$$

where L is the length of the spreading code and Fc is the chipping rate.

and $$Bss = \frac{1}{Tc} = Fc \tag{7}$$

Therefore, $$Tp = \frac{L}{R^2 \, Fc^3 \left(\frac{S}{N}\right)_{out}} \tag{8}$$

Expression (8) above shows the relationship between the required length of the location burst, Tp, and the desired timing jitter on the TOA estimate for a given output signal-to-noise ratio, SNR. This is important because we are interested in the minimum length of a burst transmission in order to achieve as high a capacity as possible.

From this expression it appears that the duration of the location burst is inversely proportional to the cube of the chipping rate. This is true for a constant SNR but as the chipping rate is increased, the bandwidth is increased and thus the transmitted power must be increased in order to maintain the SNR.

The longer the code, L, the longer the location burst. The longer the code, the fewer sequences, or epochs[2], in a given time, hence less averaging is possible. Thus, as far as capacity is concerned, it is desirable to have as short a code as practical.

From the definition of processing gain,

$$(S/N)_{out} = PG(S/N)_{in} \tag{9}$$

thus,

$$(S/N)_{in} = \frac{(S/N)_{out}}{PG} \tag{10}$$

from (6) and (7)[3]

$$PG = BssTs = L \tag{11}$$

therefore,

$$(S/N)_{in} = \frac{(S/N)_{out}}{L} \tag{12}$$

Expression (8) can be rewritten as:

$$Tp = \frac{1}{R^2 Fc^3 (S/N)_{in}} \tag{13}$$

Now

$$(S/N)_{in} = \frac{\mathrm{Pr}}{No} \times \frac{1}{Bss} = \frac{\mathrm{Pr}}{No} \times \frac{1}{Fc} \tag{14}$$

where Pr is the received power and No is the noise power per unit bandwidth.

Therefore,

$$Tp = \frac{1}{R^2 Fc^2 (\mathrm{Pr}/No)} \tag{15}$$

Expression (15) shows that the duration of the location pulse is inversely proportional to the square of the chipping rate or bandwidth. Hence the capacity of the system is proportional to the square of the bandwidth.

---

[2]One code sequence is known as an "epoch".
[3]This is true if each code sequence, or epoch, is correlated for the determination of TOA. It is also true if the data encoded into the PN sequence is one data symbol per sequence, but if more data bits per sequence are encoded, then the effective PG for the reception of the data is decreased.

2. Relationship between Data Capacity and Bandwidth

A code sequence, length L, is known as an epoch. Data is encoded into the location burst by phase invertions of the PN sequence. Usually one data symbol is encoded per epoch, which results in the processing gain being equal to the code length, L[4]. In expression (2), M, the number of independent samples, is therefore also the number of epochs or data symbols

Now the length of a pulse, $Tp = M.Ts$ where Ts is the duration of a single spread spectrum sequence.

Therefore, from (6)

$$M = \frac{TpFc}{L} \tag{16}$$

Now, from (15),

$$Tp \propto 1/B^2 \tag{17}$$

and, from (7)

$$Fc \propto B \tag{18}$$

Therefore, substituting (17) and (18) into (16)

$$M \propto \frac{1}{BL} \tag{19}$$

In unit time, there are $\frac{1}{Tp}$ pulses. Therefore, from (17) and (19), the number of data symbols, S, in unit time are:

$$S = \frac{M}{Tp} \propto \frac{B}{L} \tag{20}$$

Thus, if the bandwidth is doubled, the number of data symbols per second are also doubled, assuming the code length, L, is constant.

[4]If more than one data symbol is encoded per epoch, then the processing gain is reduced, e.g. if four data symbols are encoded per epoch, then the processing gain will be L/4.

3. Processing Gain and Jamming Margin

The longer the code, L, the longer the location burst. The longer the code, the fewer epochs in a given time, hence less averaging is possible. Thus, as far as capacity is concerned, it is desirable to have as short a code as practical. Short codes, however, equate to lower processing gain (PG) and smaller jamming margin. This is shown as follows:

Now, $$(S/N)_{in} = \frac{(S/N)_{out}}{L} \qquad (12)$$

Expression(12) has assumed that the PG is equal to L . From expression (12) it can be seen that the required input SNR is related to the inverse of the code length. Therefore, as the code length is reduced, the higher the required input SNR.

For example, with a code length of 255, and a required output SNR of 10dB, the required input SNR is 10.log(10/255) = -14dB. This means that the power of any interfering or jamming signal, within the spread bandwidth, spread spectrum or narrow band, needs to be 14dB higher than the wanted signal in order to desensitize the wanted signal. This value of 14dB is known as the **Jamming Margin (JM)** and is more often seen in the following familiar expression of:

$$JM = 10.\log PG - (S/N)_{out} \quad \text{(JM and S/N in dB)} \qquad (21)$$

4. Near-Far-Ratio

The Jamming Margin can be expressed as:

$$\frac{P_{rj}}{P_r} = \frac{PG}{(S/N)_{out}} \quad \text{where Prj is the power of the received jamming signal} \tag{22}$$

In the Hata formula[5] the propagation loss due to distance is:

(44.9 - 6.55 log h_b) log D where h_b is the base station antenna height (m) and D is the distance (kms.)

Thus for a 100ft (30m) mast, the distance loss is 35.22 log D and for a 300 ft mast, the distance loss is 31.8 log D. Therefore, for a 100 ft mast, the propagation loss[6] is proportional to $D^{3.5}$.

Hence
$$\frac{PG}{(S/N)_{out}} = \frac{D^{3.5}}{D_j^{3.5}} \tag{23}$$

Therefore
$$\frac{D_j}{D} = \sqrt[3.5]{\frac{PG}{(S/N)_{out}}} = NFR \quad \text{(this is known as the Near-Far Ratio)} \tag{24}$$

---

[5]The accepted formulas for the prediction of propagation loss in an urban environment are those in CCIR Recommendation 370-1 which are based on the Okumura prediction method (Y. Okumura et al., "Field strength and its variability in in UHF and VHF land mobile service",Rev. Elect. Commun. Lab., vol 16,1968). An empirical formula for propagation loss, derived from Okumura's report has been produced by Hata ("Empirical Formula for Propagation Loss in Land Mobile Services", IEEE Trans. on Veh. Tech., vol VT-29, No.3 1980). This formula has become standard in planning for land mobile systems.

[6]Often the propagation loss due to distance is taken as D^4 (as is used in the Egli propagation formula).

5. Range

Expression (12)shows that the input SNR was inversely related to the PG or L, the length of the spreading code. The input SNR is related to the transmitted power and the transmission propagation or path loss. If the system requires a higher input SNR than another system, then, for equal transmission power, one system will have a greater range than the other.

From the Hata formula for a 100 ft mast, the propagation loss is proportional to $D^{3.5}$.

Now
$$(S/N)_{in} = \frac{\mathrm{Pr}}{N} \propto \frac{Pt}{N.D^{3.5}} \qquad (25)$$

where Pr and Pt are the received and transmitted powers respectively and N is the received noise power.

As $N \propto Bss$,
$$(S/N)_{in} \propto \frac{1}{Bss.D^{3.5}} \qquad (26)$$

Combining with (10)and (7)
$$(S/N)_{out} \propto \frac{PG}{Fc.D^{3.5}} \qquad (27)$$

Hence
$$D \propto \sqrt[3.5]{\frac{PG}{Fc}} \qquad (28)$$

6. Multipath Error

The error, R, as given in Section 1 of this paper, is the time-of-arrival error due to noise at the receiver. In practice, the correlated epochs will have a time-varying component due to the combined effects of vehicle movement and signal multipath. This is known as "variable multipath error". There is also a slow time-varying component that is known as "multipath bias".

Due to the summation of several multipath signals, the received signal is complex, however, over a short period, the contribution of each component can be expressed as:

$$Em = A\cos(2\pi f_d t + \phi) \qquad (29)$$

where Em is the epoch error due to multipath
A is amplitude of error
f_d is Doppler frequency
ϕ is phase constant

If averaged over time T the mean error is:

$$\overline{Em} = \frac{1}{T}\int_0^T \cos(2\pi f_d t + \phi)\,dt$$

$$= \frac{A}{2\pi f_d T}\sin(2\pi f_d T) \qquad (30)$$

For the amplitude of multipath error to be reduced, $2\pi f_d T \geq 1$ (31)

Hence $$f_d \geq \frac{1}{2\pi T} \qquad (32)$$

Now $$f_d = V/\lambda \qquad (33)$$

where V is the speed of the vehicle
λ is the received signal wavelength

From (32) and (33), $$V \geq \frac{\lambda}{2\pi T} \qquad (34)$$

There is therefore a minimum vehicle speed below which there is no reduction of the variable multipath error by averaging. Therefore, TOA errors from stationary or slow moving vehicles can be expected to be higher than those from moving vehicles.

The integration time, T in expression (34), is the duration of the location pulse, Tp, from expression (15). The longer the duration of Tp, the slower the required speed of the vehicle in order to reduce the variable multipath error by averaging. Table 1 shows the relationship.

Table 1 - Minimum vehicle speed for improvement of variable multipath errors	
Tp	**Vmin**
1ms	117.8 mph
4ms	29.5 mph
10ms	11.8 mph
50ms	2.4 mph
100ms	1.2 mph
1sec	0.12 mph

From Table 1, it can be seen that for location pulse durations under 4 ms, no improvement of the variable multipath error will result for vehicles travelling at 30 mph or less. However, if the short duration of the pulse, Tp, is the result of a high chipping frequency, as per expression (15), then the ability to resolve the multipath component is better. This is explained as follows:

A standard direct sequence receiver cannot resolve delayed multipath receptions of less than one chip time. The ability to resolve the direct and delayed signals is also dependent upon their relative amplitudes.

Assuming that the peak multipath error A=Tc, then, from (30) and (34),

$$\hat{E}m = Tc\left(V_{\min}/V\right) \qquad \text{for } V \geq V\min$$

$$\hat{E}m = Tc \qquad \text{for } V < V\min \qquad (35)$$

It also follows that the multipath bias error, Emb, which has a long time constant, will be equal to Tc, i.e.

$$\hat{E}_{mb} = Tc \qquad (36)$$

Hence the shorter the chip period, the lower the multipath bias error.

7. Effect of Bandwidth (Chipping Rate) and Code Length

Consider two systems, A and B, whose basic parameters are the same except for Fc, the chipping rate, and L, the code length..

From (15)
$$\frac{T_{pA}}{T_{pB}} = \left(\frac{F_{cB}}{F_{cA}}\right)^2 \tag{37}$$

From (20), the number of data symbols per unit time,

$$\frac{S_A}{S_B} = \frac{F_{cA} L_B}{F_{cB} L_A} \tag{38}$$

From (28), the range
$$\frac{D_A}{D_B} = \sqrt[3.5]{\frac{F_{cB} L_A}{F_{cA} L_B}} \tag{39}$$

From (35) the variable multipath error,

$$\frac{\hat{E}_{mA}}{\hat{E}_{mB}} = \frac{T_{cA}}{T_{cB}} \times \frac{T_{pB}}{T_{pA}} = \frac{F_{cA}}{F_{cB}} \tag{40}$$

and from (36), the bias error,

$$\frac{\hat{E}_{mbA}}{\hat{E}_{mbB}} = \frac{T_{cA}}{T_{cB}} = \frac{F_{cB}}{F_{cA}} \tag{41}$$

7.1. Example 1, double bandwidth and constant code length.

Let the chipping rates of A and B be 2 Mcps and 1 Mcp respectively, and let $L_A = L_B$.

Duration of location pulse	$\frac{T_{pA}}{T_{pB}} = \frac{1}{4}$
Data symbols per unit time	$\frac{S_A}{S_B} = 2$
Range	$\frac{D_A}{D_B} = 0.82$
Variable multipath error	$\frac{\hat{E}_{mA}}{\hat{E}_{mB}} = \frac{2}{1}$
Multipath bias error	$\frac{\hat{E}_{mbA}}{\hat{E}_{mbB}} = \frac{1}{2}$

Therefore, the effect of doubling the bandwidth is to:

1) increase the system capacity by a factor of four,
2) increase the data capacity by a factor of two,
2) reduce the range by 0.82.
3) double the variable multipath error,

and 4) halve the multipath bias error.

The significant gain in system capacity is offset by a reduction in range. The effect on multipath error tends to cancel out if both stationary and moving vehicles are considered. The 0.82 reduction in range equates to reducing the area of a cluster by 0.66.

7.2. Example 2, Double bandwidth and code length..

Let the chipping rates of A and B be 2 Mcps and 1 Mcp respectively, and let $L_A = L_B$.

Duration of location pulse $\quad \dfrac{T_{pA}}{T_{pB}} = \dfrac{1}{4}$

Data symbols per unit time $\quad \dfrac{S_A}{S_B} = 1$

Range $\quad \dfrac{D_A}{D_B} = 1$

Variable multipath error $\quad \dfrac{\hat{E}_{mA}}{\hat{E}_{mB}} = \dfrac{2}{1}$

Multipath bias error $\quad \dfrac{\hat{E}_{mbA}}{\hat{E}_{mbB}} = \dfrac{1}{2}$

Therefore, the effect of doubling the bandwidth is to:

1) increase the system capacity by a factor of four,
2) maintain the same data capacity,
2) maintain the same range,
3) double the variable multipath error,
and 4) halve the multipath bias error.

7.3. Discussion of Example Results

In comparing the results of Example 2 with Example 1 it can be seen that the range is maintained at the expense of data capacity. From these examples it can be seen that the two requirements of location capacity and data capacity are somewhat in opposition. For location efficiency, If the bandwidth is increased, by the use of a faster chipping rate, it is better to also increase the code length so as to maintain the range. This results, however, in no gain to the data capacity of the location burst. If the location burst is used for data transmissions, then it is better to keep the code length, L, constant. But, as the chipping rate is increased, in order to increase the location capacity, the range reduces. Thus, if considering data capacity, it is better to have as short a code length as possible which results in poor processing gain and poor range. For best location capacity it is better to have as long a code length as possible[7].

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[7] As the code length is increased, the frequency accuracy demand becomes higher and also the time to synchronize becomes longer. Therefore there is a practical limit to the code length of about 1023.

APPENDIX A

DERIVATION OF TOA ERROR

The received spread spectrum waveform is correlated in the spread spectrum receiver. Figure 1 represents a typical correlation peak out of the correlator; this diagram will serve as a basis for deriving the desired timing error relationship[1] .

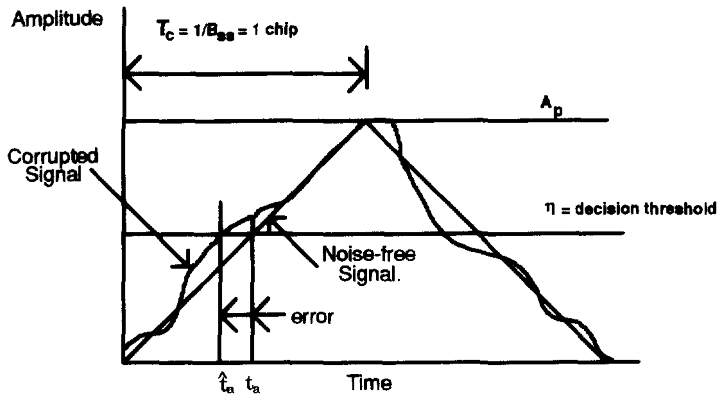

Figure 1. Correlated location pulse.

At $\hat{t}_a$, the time at which the correlator output waveform crosses the decision threshold, the signal can be considered as the sum of the noise-free signal term and a noise term, described as

$$A(\hat{t}_a) + n(\hat{t}_a) = \eta \tag{1}$$

Assuming that the rising edge of the correlator output pulse is approximately linear, the amplitude of the noise-free signal at the decision threshold can be written as

$$A(\hat{t}_a) = \frac{\hat{t}_a}{T_c} A_p \tag{2}$$

From (1) and (2),

$$\hat{t}_a = \left[\eta - n(\hat{t}_a)\right]\frac{T_c}{A_p} \tag{3}$$

Now

$$A_p = \sqrt{S_{out}} \tag{4}$$

where S_{out} is the peak signal power at the correlator output.

[1] This derivation is similar to that given in Comments of Pinpoint Communications, Inc., June 29, 1994, Exhibit A.

Substituting(4) into (3), $$\hat{t}_a = \left[\frac{\eta - n(\hat{t}_a)}{\sqrt{S_{out}}}\right] T_c \tag{5}$$

The variance of the random variable $\hat{t}_a$ gives the tracking jitter on the time of arrival estimate.

$$VAR(\hat{t}_a) = \sigma_c^2 = E\left[(\hat{t}_a)^2\right] - \left[E(\hat{t}_a)\right]^2 \tag{6}$$

Now $E(\hat{t}_a)$ is the mean value of the noise, and assuming that the noise is zero mean,

$$E(\hat{t}_a) = t_a \tag{7}$$

From (5) $$E\left[\hat{t}_a^{\,2}\right] = \frac{T_c^{\,2}}{S_{out}}\left[\eta^2 - 2\eta E(n(\hat{t}_a)) + E(n^2(\hat{t}_a))\right] \tag{8}$$

Now $E(n(\hat{t}_a))$ is zero (assuming zero mean noise), thus,

$$E\left[\hat{t}_a\right] = \frac{T_c^{\,2}\eta^2}{S_{out}} + \frac{T_c^{\,2}}{S_{out}} E(n^2(\hat{t}_a)) \tag{9}$$

The noise power at the correlator output is the variance of $n(\hat{t}_a)$, hence

$$No = E\left[n(\hat{t}_2)\right]^2 + E\left[n^2(\hat{t}_a)\right]$$

$$= E\left[n^2(\hat{t}_2)\right] \tag{10}$$

Now $$\frac{T_c\eta}{A_p} = t_a$$

Hence $$\frac{T_c^{\,2} n^2}{S_{out}} = t_a^{\,2} \tag{11}$$

Substituting (10) and (11) into (9)

$$E\left[\hat{t}_a\right] = t_a^{\,2} + \frac{T_c^{\,2}}{S_{out}} No \tag{12}$$

Substituting (12) and (7) into (6)

$$\sigma_t^{\,2} = \frac{T_c^{\,2} No}{S_{out}} \tag{13}$$

Defining $B_{ss} = \dfrac{1}{T_c}$,

$$\sigma_t^2 = \frac{1}{B^2 ss\left(\frac{S}{N}\right)_{out}} \tag{14}$$

Out of Band Transmissions for LMS Allocation
Recommended Specification

Summary

The "Emission Limitations", given in Section 21.106 of the FCC Regulations, serves to set the limits on high speed digital data and as such is directly applicable to the out of band emissions for spread spectrum signals. So as to directly relate the specification to maximum permitted transmitted power output, it is proposed that the specification be slightly amended to read as follows:

"For LMS systems, operating in the 902 -928 MHz band, in any 4 kHz band, the center frequency of which is removed from the center of the authorized band by more than 50 percent up to and including 250 percent of the authorized bandwidth, as specified by the following equation but in no case less than 50 dB:
$A = 35 + 0.8 (P - 50) + 10 \log B$ (attenuation greater than 80 dB is not required)
where A = attenuation (in decibels) below the maximum permitted mean output power level,
P = percent removed from the carrier frequency,
B = authorized bandwidth in megahertz."

1. Introduction

A specification of out of band transmissions must refer to the allocated band and take into account differing transmitted powers. For example, it is not sufficient to only specify the relative level of the sidelobes to the peak level[1], with respect to the spectral envelope as this does not relate to the allocated bandwidth or to the actual transmitted power.

2. Previously Suggested Specifications

Teletrac[2] and Pinpoint have suggested a specification of 99% of transmitted power within the allocated bandwidth. This specification does not restrict the emissions with respect to frequency separation and is not related to transmitted power level..

MobileVision[3] suggested a specification similar to the "Transmitter Sideband Spectrum" standard as given in EIA/TIA-316-C[4]. The suggested specification was:

i) The envelope spectrum shall be attenuated at least 35 dB from the peak of the signal at any frequency spaced from the center frequency by more than 50% of the authorized bandwidth.

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[1]SBMS suggested -20 dB for first sidelobe with each following sidelobe progressively reduced by 10 dB. SBMS Comments, June 29, 1993, p.24.
[2]Teletrac Comments, June 29, 1993, p.50.
[3]MobileVision Comments, June 29, 1993, Annex A, p.20.
[4]Minimum Standards for Portable/Personal Radio Transmitters, etc., 25-1000 MHz.

ii) The envelope spectrum shall be attenuated at least 50 dB from the peak of the signal at any frequency spaced from the center frequency by more than 100% of the authorized bandwidth.

For the MobileVision specification to be compatible with the "99%" rule, suggested by Teletrac and Pinpoint, rule (i) would need to be changed from 35 dB to 20 dB and rule (ii) deleted.

MobileVision's suggested specification does limit the out of band transmissions with respect to the allocated bandwidth and with respect to ensuring that the emissions decrease with separation, it does not, however, relate to the transmitted power. For example, a mobile transmitting 1W would still have to meet the same spectral envelope even though the out of band emissions would be 10 dB less than a mobile transmitting 10W.

3. Proposed Out of Band Specification (FCC Spectrum Envelope).

The FCC spectrum envelope, given in Section 21.106, serves to set the limits on high speed digital data bandwidth and as such is directly applicable to the out of band emissions for spread spectrum signals. The rule reads as follows:

"For operating frequencies below 15 GHz, in any 4 kHz band, the center frequency of which is removed from the assigned frequency by more than 50 percent up to and including 250 percent of the authorized bandwidth, as specified by the following equation but in no case less than 50 dB:

$A = 35 + 0.8 (P - 50) + 10 \log B$ (attenuation greater than 80 dB is not required)

where A = attenuation (in decibels) below the mean output power level,
P = percent removed from the carrier frequency,
B = authorized bandwidth in megahertz."

It is proposed that the above specification be adopted for application to the LMS band with the following changes:
Replace the phrase "assigned frequency", with "center of the authorized band":
and redefine A as

"A = attenuation (in decibels) below the maximum permitted mean output power level," The revised, and hence proposed specification therefore reads as follows:

"For LMS systems, operating in the 902-928 MHz band, in any 4 kHz band, the center frequency of which is removed from the center of the authorized band by more than 50 percent up to and including 250 percent of the authorized bandwidth, as specified by the following equation but in no case less than 50 dB:

$A = 35 + 0.8 (P - 50) + 10 \log B$ (attenuation greater than 80 dB is not required)

where A = attenuation (in decibels) below the maximum permitted mean output power level,

P = percent removed from the carrier frequency,

B = authorized bandwidth in megahertz."

This specification has the advantages of being related to the maximum permitted output power level, the allocated bandwidth and the frequency separation.

4. Application of Proposed Specification.

It is worthwhile carrying out an example so as to realize the specification limits.

Let the allocated bandwidth, B = 8 MHz

Let the frequency separation from the center of the authorized band be Δf.

Thus,	Δf = 4 MHz	P = 50	A = 50 (calculates to 44, but 50 is minimum)
	Δf = 6 MHz	P= 75	A = 64
	Δf = 8 MHz	P = 100	A = 80 (calculates to 84 but 80 is maximum)

Thus, if the maximum permitted mean transmitted power is 10W (40 dBm), then the out of band emissions must be:

Δf	max. emission (in 4 kHz B/W)
4 MHz	-10 dBm
6 MHz	-35 dBm
8 MHz	-60 dBm

To convert this to the spectral envelope requirement, it is necessary to calculate the power contained in a 4 kHz band. Thus it is necessary to account for the noise bandwidth of the modulation.

The relationship between modulation and the noise bandwidth is:

Modulation	Noise Bandwidth per bit rate
BPSK	1.00
QPSK	0.5
MSK	0.62

Therefore the noise bandwidth of a BPSK, 2 Mcps spread spectrum signal is 2 MHz.

Assuming a 2 Mcps, BPSK spread spectrum signal, the power in a 4 kHz bandwidth is:

$$10 \log 4/2000 = -27 \text{ dB}$$

The spectral envelope requirement is therefore:

Δf	Attenuation below peak
4 MHz	-23 dB
6 MHz	-37 dB
8 MHz	-53 dB

As the chipping rate and modulation type is changed, the spectral envelope will vary but the out of band emissions, measured in a 4 kHz bandwidth, remain constant.

G K Smith
3/24/94
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JOINT DECLARATION OF
ANTHONY J. SPADAFORA AND GRAHAM K. SMITH

We, Anthony J. Spadafora and Graham K. Smith, are Vice President, Technology and Director, Systems Research, respectively, for METS, Inc., the general partner of MobileVision, L.P. We have prepared the foregoing technical annexes accompanying these Further Reply Comments of MobileVision, L.P. in response to the Further Comments concerning the ex parte submissions filed in PR Docket No. 93-61, Amendment of Part 90 of the Commission's Rules to Adopt Regulations for Automatic Vehicle Monitoring Systems. We declare under penalty of perjury that the foregoing technical annexes, to the best of our knowledge, are true and correct.

Anthony J. Spadafora

Graham K. Smith

Date: March 28, 1994

CERTIFICATE OF SERVICE

I, Lila A. Mitkiewicz, hereby certify that copies of the foregoing **Further Reply Comments of MobileVision, L.P.** were forwarded this 29th day of March, 1994 by U.S. first-class mail to the following individuals:

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